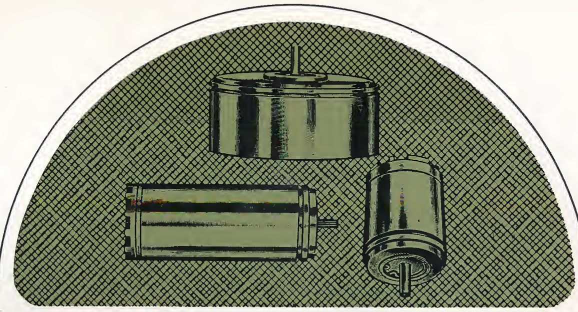


SOLVERE TECH-TIPS

ISSUE NUMBER TWO

Tech-Tips is a technical publication prepared by Solvere, Inc., Santa Ana, California. It is designed to provide information on precision servo components such as resolvers, motor tachs, hysteresis synchronous motors and integrating motor generators. We invite inquiries on your most stringent requirements within these product areas.

RESOLVER PARAMETERS AFFECTED BY TEMPERATURE CHANGES AND THE TECHNIQUES FOR TEMPERATURE COMPENSATION.

The basic change which takes place in a resolver over temperature swings is a change in DC resistance of the primary and secondary windings.

The DC resistance of copper magnet wire will vary directly with temperature at a rate of 0.4%/°C.

With variations in DC resistance we expect to see variations in those resolver parameters which are directly related to DC resistance (Rdc). The two parameters which are affected are phase shift, secondary to primary and the transformation ratio.

The phase shift through a resolver is determined by the DC resistance of the primary and the reactive portion of the primary impedance (X_L).

$$\text{Phase shift} = \text{Tan}^{-1}\frac{\text{Rdc}}{X_L}$$

For purposes of estimating phase shift change with temperature, we assume that X_L remains constant (not quite true), therefore, the phase shift will increase with increasing temperature.

Example:

If phase shift at 25°C = 8°, what will phase shift be @ 50°C?

At 25°C

$8° = \tan^{-1} .14$

At 50°C

$$\text{Phase shift} = \tan^{-1}\left(\frac{\Delta t \times .4\% \times \text{Rdc}}{X_L}\right) + \frac{\text{Rdc}}{X_L}$$

$$\text{Phase shift} = \tan^{-1}\left(\frac{25 \times .4 \times .14}{100}\right) + .14$$

$$= \text{Tan}^{-1} .014 + .14$$

$$= \tan^{-1} .154$$

Phase shift = 8.75°

The in-phase transformation ratio of a resolver is a function of the output of the unit times the cosine of the phase shift angle. Therefore, it can be seen that as temperature increases, the phase shift increases, and the transformation ratio will decrease.

Using the same example as above, assume output at 25°C = 1.0

At 25°C

$\text{TR} = \text{Eout} \times \cos\Theta$

$\text{TR} = 1.0 \times .990$

$\therefore \text{TR} = .990$

At 50°

$\text{TR} = \text{Eout} \times \cos\Theta$

$= 1.0 \times \cos 8.75°$

$.988$

In many case variations in phase shift and transformation ratio can not be tolerated over temperature extremes. Since both of these parameters vary as a function of DC resistance it is possible to compensate for this resistance change.

A thermistor in the primary winding (buried inside the resolver), will produce in effect, a constant DC resistance over the temperature range. The proper selection and shunting of the thermistor is important and Solvere has applied this technique to many units.

Thermistor compensation in most cases results in a higher initial phase shift which may be permissible but if not can be reduced to zero by means of a trimming capacitor across the output winding.

As an example of what can be done with temperature compensation a typical size 8 resolver can be held to: TR ± .3% and Phase shift ± 1° over temperatures −54°C to +100°C.

In addition to variations caused by DC resistance change, there can also be variations caused by mechanical changes. These are primarily due to the squeezing effects because of differential expansions.

Because of the unpredictable results of such squeezing it is difficult to compensate. However, through careful selection of materials and skillful design, it is possible to minimize the problems in this area.

The thermistor method of temperature compensation is used in systems where frequency variation is not a problem and where some stability over temperature ranges is needed.

If frequency variations as well as temperature variations are present, the booster amplifier method, with compensator winding resolver is the approach normally taken. This method is described in detail in another section of this issue of Tech-Tips.

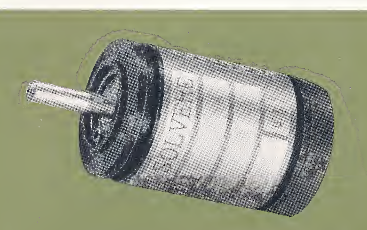
LINEAR POT HAS 4′ ACCURACY THROUGH ±70° ANGULAR TRAVEL

This new linear potentiometer has a 4′ accuracy over a range of ±70° angular travel. Designated the *Model 8SS 20,* it is housed in a size 8 case measuring .75″ (dia.) x 1.175″. Weight is 1.7 ounces.

Typical specifications for the Model 8SS 20 based on 26 V, 400 cycle input are:

Input current:	26 ma
Phase shift:	12.5°
^{Z}RO:	240 + 934
^{Z}ss:	666 + j270
Voltage gradient:	510 mv/degree

The Model 8SS 20 is designed to meet the environmental requirements of MIL-E-5272C. Temperature range is −55° to +125°C.



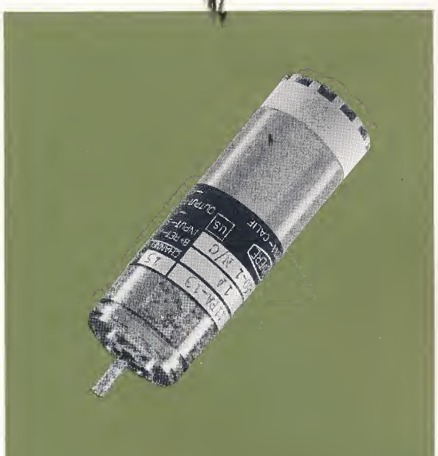
INDUCTION POTENTIOMETER (8SS20): *infinite resolution; brushless; wide temperature range; excellent shock, vibration and temperature characteristics; high sensitivity; can be modified for space environments (hard vacuum, radiation).*

INTEGRAL RESOLVER-AMPLIFIER COMBINATION HAS 0.1% RESOLVER ACCURACY. SAVES SPACE, WEIGHT, TRIM TIME

Solvere's new resolver-amplifier combination is housed in a size 11 case measuring only 1.062″ (dia.) x 3″ (including terminals). Weight is 7 ounces. By combining both the amplifier and the resolver in a single package, Solvere has accomplished a savings in size, weight, test time, trim time and mounting time. An additional benefit is improved reliability through the elimination of connectors and hardware between the resolver and amplifier.

The Model 11PA13 is trimmed to 0.05% TR and 1 minute phase shift at room temperature. It will hold TR to 0.1% and phase shift to 6 minutes over a temperature range of −54°C to 100°C.

Standard amplifier supply is 50 VDC with other voltage levels available. Input impedance to amplifier is over 500,000 ohms.



RESOLVER-AMPLIFIER (11PA-13): *Integral package for reduced wiring problems; less total volume; single source responsibility. TR and phase shift held over wide temperature range. Series summing (parallel summing also available in same package size.)*

Here is the second in a series of technical discussions on precision resolvers. The first issue contained a brief explanation of resolvers, how resolvers are used and resolver inputs. Subsequent issues will deal with resolver windings, frequency response, application factors, etc., as well as similar discussions on other servo components.

Equivalent Circuit

The resolver equivalent circuit is useful because variations in operating conditions act primarily to change the complex transformation ratio of maximum secondary voltage to primary voltage without appreciably affecting the essentially sinusoidal character of the output voltage variation with shaft angle. Furthermore, for balanced loads and low primary source impedance, the input and output impedances are independent of shaft position. Therefore, useful information can be obtained from the conventional transformer-type equivalent circuit corresponding to the four-terminal network formed by one primary and one secondary winding at maximum coupling. Figure 1 shows the basic uncompensated resolver equivalent circuit. In this position, neither of the windings couple the remaining two so those that are not shown can be ignored.

Often special data are required for special resolver applications. If a prototype unit is available, its equivalent circuit can be measured and the desired characteristics determined by calculation. When taking the measurements, a primary and secondary winding must be aligned for maximum mutual coupling. Because of the small size of these components, their leakage impedances are high and bridge-type measurements give more satisfactory results.

The equivalent-circuit constants can also be determined from manufacturer's data.

Transformation Ratio and Phase-Angle Errors

As mentioned previously, frequency, temperature, secondary loading, and supply voltage changes affect the equivalent-circuit constants and cause corresponding changes in the transformation ratio and phase angle of a resolver. Circuit reactances vary directly with changes in supply frequency; winding copper resistance responds to ambient temperature variations; and changes in the iron core permeability with flux density cause jX_m to change with voltage. The leakage reactance and resistance of the resolver windings cause a drop in output voltage analogous to the regulation characteristics of a transformer.

The equivalent circuit also shows that the resolver nominal phase shift differs from zero, and in the smaller sizes can approach 4 or 5°.

Standardizing and temperature-compensating networks, together with rigid specifications of circuit impedances and loading, permit resolvers to be used for medium-accuracy computer work without the associated electronic equipment frequently needed in the more accurate applications. Here the equivalent circuit is useful in establishing circuit-component tolerances and the permissible variations in operating tolerances.

Standardization and compensation networks are usually incorporated in booster amplifiers. Boosters are high-gain amplifiers with adjustable negative feedback elements, easily adapted for standardizing the over-all gain and phase shift of the booster plus resolver combination. Usually the over-all phase shift is held accurately to zero degrees, with a transformation ratio of 1:1. Elements to compensate for variations in line voltage, frequency, and ambient temperature are usually included in the booster feedback system.

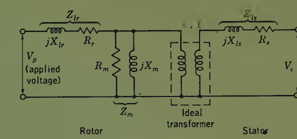


Fig. 1. Equivalent circuit of uncompensated resolver.

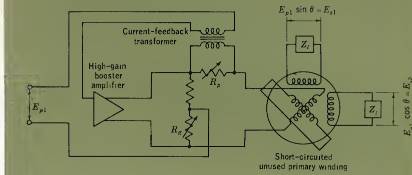


Fig. 2. Network-type booster compensation circuit. Z_l = standard secondary load; R_p, R_g = phase and gain attenuators; θ = rotor-shaft angle.

Boosters are also used as isolators to avoid loading computing devices that would normally feed the low-impedance primary windings of the resolver directly, and likewise to avoid resolver loading.

Two typical booster arrangements are shown in Figs. 2 and 3. These show excitation applied to only one resolver primary. Similar circuits would be required for exciting the other primary. Thus, a resolver with two input voltages requires two amplifier channels. These are frequently built on one chassis and sold as twin boosters. Since normally the secondaries are lightly loaded, no special adjustments are necessary. The effect of the load is included in the primary standardization.

The circuit of Fig. 2 uses two adjustable feedback loops. In one of them R_p is set to be proportional to R_r, the resolver primary resistance. The feed-back voltage is then proportional to the drop across R_r and is scaled by the current feedback amplifier so that the voltage across R_r is effectively canceled by the output of the amplifier. The resolver appears to have no primary resistance and the over-all phase shift is zero. The over-all gain is adjusted to a standard value by means of resistor R_g in the voltage feedback loop. Resistance R_p (proportional to R_r) is made temperature-sensitive so that proper compensation takes place over a wide range of ambient temperatures. The critical resistors in the compensation and standardization loops are located in the end cap

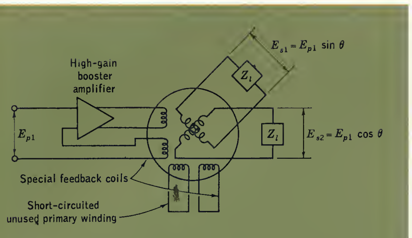


Fig. 3. Booster compensation system using special feedback windings. Z_l = standard secondary load; θ = rotor-shaft angle.

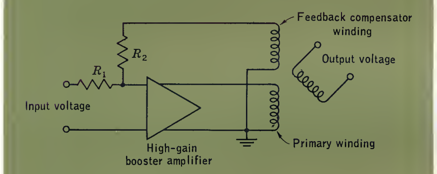


Fig. 4. Simplified schematic of compensator winding resolver-amplifier assembly.

of the resolver so that, when necessary, the booster amplifier can be replaced without subsequent circuit adjustment.

Booster System With Compensator Winding Resolver

The booster technique shown in Fig. 3 is most commonly used. Special feedback coils are wound in intimate proximity to the resolver primary windings so that the coefficient of coupling is practically unity. The output voltage from this winding is very nearly in phase with the rotor output voltage. It remains in phase despite changes in temperature and frequency, since the effects of primary copper loss are reflected equally in both the feedback and rotor windings. Resolvers of this type have their primaries wound on the stator so that extra slip rings are not required for the feedback windings.

To analyze the performance of this system, refer to the schematic of the amplifier-resolver assembly shown in Fig. 4. If the booster gain is very high, the feedback compensator winding develops a voltage that is almost equal to R_1/R_2 times the input voltage. And since the feedback winding is intimately coupled with the primary winding, its output voltage is essentially a measure of primary flux. Then the feedback amplifier delivers a primary voltage of the correct magnitude and phase to generate a flux field corresponding to the voltage applied at the input to the booster.

Thus assuming perfect coupling between the compensator and primary windings, the total flux field generated by the primary corresponds exactly to the applied voltage, thereby preventing phase shift caused by primary resistance and its corresponding temperature error. Although this compensating technique is sufficiently accurate to be widely used, a detailed study points up certain errors.

Figure 5 shows an elementary schematic of the coil arrangements in a compensator winding resolver, with maximum coupling between the primary and secondary windings. This reveals the various leakage and mutual fluxes occurring in a resolver. From this schematic it is possible to formulate an equivalent electrical circuit in which the individual flux linkages are replaced by circuit reactances (Fig. 6). On the basis of this equivalent circuit it is possible to determine the accuracy of the compensator winding method of resolver compensation.

Assume that both the secondary and compensator windings are unloaded. Also, because of the feedback circuit and the high-gain amplifier, it can be assumed that the compensator voltage is identical to the input voltage in phase, magnitude (except for a proportionality factor), and waveshape.

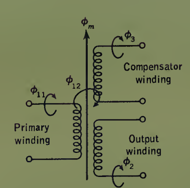


Fig. 5. Flux linkage pattern in compensator winding resolver. ϕ_{11} = flux common only to primary winding; ϕ_{12} = flux common only to primary and compensator windings; ϕ_2 = flux common only to the output or secondary winding; ϕ_3 = flux common only to compensator winding; ϕ_m = flux common to all three windings; $\phi_{11} + \phi_{12} + \phi_m$ = total flux created by excitation of the primary winding.

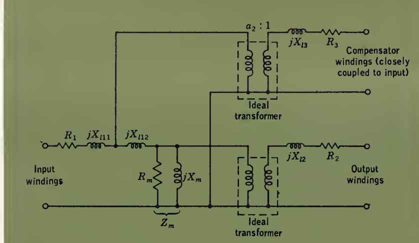


Fig. 6. Equivalent circuit for compensator winding resolver.

To determine the output voltage from the resolver secondary winding (at maximum coupling) it is necessary to work backward from the compensator output. Thus from the equivalent circuit,

$$K = \frac{V_{\text{output winding}}}{V_{\text{compensator}}} = \frac{a_2}{a_1} \frac{Z_m}{Z_m + jX_{l12}} \tag{5}$$

And if the resolver has a negligible core-loss factor so that Z_m equals jX_m, then

$$K = \frac{a_2}{a_1} \frac{X_m}{X_m + X_{l12}} \tag{6}$$

Equation 6 shows that a zero phase shift has been achieved, which, within the limits of the assumed equivalent circuit, is independent of temperature and frequency. However, at very high frequencies, stray capacities modify this expression.

Even assuming negligible core loss, a source of error results from the different nature of the X_m *and* X_{l12} reactances. Because the flux corresponding to X_m crosses a relatively small airgap (measured in thousandths of an inch) and has a substantial part of its reluctance in the nonlinear iron, while the flux corresponding to X_{l12} is essentially generated across the leakage airgaps, the ratio K will vary with the applied voltage in a similar manner to the variation of iron permeability with flux density. However, by using high-nickel alloy steels, it is possible to minimize this effect and limit it to a few hundredths of 1 per cent. (The compensator winding method of phase compensation nullifies this effect in relation to the phase of the output voltage.)

Now consider the effect when the core loss is not assumed negligible. The phase shift can be determined as the phase angle of

$$\frac{Z_m}{Z_m + jX_{l12}}$$

Establish an expression for Z_m. Let Z_m consist of jX_m, in parallel with R_m equals QX_m. Then

$$Z_m = \frac{jX_m Q}{Q + j} \tag{7}$$

Let θ_m equal the phase angle of Z_m. Then θ_m can be determined from

$$\theta_m = \frac{1}{\pi} - \arctan\frac{1}{Q} \tag{8}$$

And the phase angle of the output θ_0 is

$$\theta_0 = -\arctan\frac{X_{l12}}{Q(X_m + X_{l12})} \tag{9}$$

Let
$$\frac{X_{l12}}{X_m} = \tau \tag{10}$$

Then the output phase angle is

$$\theta_0 = -\arctan\frac{\tau}{Q(1+\tau)} \tag{11}$$

which for small angles is very closely equal to

$$\theta_0 = -\frac{\tau}{Q(1+\tau)} \tag{12}$$

In a typical resolver, Q might equal 8 and τ equal 0.10. Then the output phase angle would be

$$\theta^\circ = \frac{0.1}{(8)(1.1)} = -0.65^\circ = -39.0'$$

Referring to the amplifier schematic of Fig. 4, it can be seen that the phase shift can be brought to zero by adding a capacitor across the feedback resistor. When properly adjusted, the feedback voltage from the compensator would lead the input by 39′, while the phase angle of the secondary output voltage would be zero.

In this circuit, the booster amplifier should have a net loop-voltage gain of 60 to 80 db. Tests of typical units indicate that the ratio of the voltage delivered by the feedback winding to that delivered by the rotor does not vary by more than about ± 0.05 per cent, and that the electrical phase shift is constant to within $1'$ or less over a temperature range from -55 to $+85°C$. When used with a suitable booster amplifier, the ratio of resolver rotor output to amplifier input has a similar stability. These results are based on no-load operation.

Using a booster has little effect on such factors as residual voltage, axis alignment, or angular accuracy. Its primary function is to preserve the transformation ratio and phase shift over the allowable range of operating conditions. Only in the case of booster-compensated units are accurate transformation ratio and phase specifications guaranteed by the manufacturer. But the user can obtain high accuracy for the boosterless types by the careful use of network standardization methods.

(continued next issue)

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